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APPROXIMATION BY A SUM OF COMPLEX EXPONENTIALS UTILIZING THE PE--ETC(U)
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APPROXIMATION BY A SUM OF COMPLEX EXPONENTIALS
UTILIZING THE PENCIL OF FUNCTION METHOD

Tapan K. Sarkar¹
Donald K. Weiner²
Vijay K. Jain³

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ABSTRACT: The approximation of a function by a sum of complex exponentials, in general, is a nonlinear optimization problem. The optimization problem, however, is linearized through the application of the pencil of function method. This non-iterative method yields the best exponential approximation for a given order of approximation. The method differs radically from the classical Wiener least squares approach in the sense that exponents calculated by the pencil of function method are directly proportional to the integrated squared error in the approximation. As the integrated squared error approaches zero, the exponents calculated by the pencil of function method approach the best least squares exponents in a continuous fashion. Among the advantages of the method are its natural insensitivity to noise in the data and explicit determination of the signal order. Examples are presented to illustrate the stability of this technique especially when noise is present in the data.

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This work has been supported by the Office of Naval Research under contract N00014-79-C-0598.

1. Dept. of Electrical Eng.
Rochester Inst. of Tech.
Rochester, NY 14623
2. Dept. of Electrical Eng.
Syracuse University
Syracuse, NY 13210
3. Department of Electrical Engineering
University of South Florida
Tampa, Florida 33620

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REPORT DOCUMENTATION PAGE		READ INSTRUCTIONS BEFORE COMPLETING FORM
1. REPORT NUMBER TR-81-8	2. GOVT ACCESSION NO. 10 11 11 11 11 11	3. RECIPIENT'S CATALOG NUMBER 11 11 11 11 11 11
4. TITLE (and Subtitle) Approximation by a Sum of Complex Exponentials Utilizing the Pencil of Function Method		5. TYPE OF REPORT & PERIOD COVERED Technical Report No. 8
		6. PERFORMING ORG. REPORT NUMBER
7. AUTHOR(s) T.K.Sarkar, D.D.Weiner and V.K.Jain		8. CONTRACT OR GRANT NUMBER(s) N00014-79-C-0598
9. PERFORMING ORGANIZATION NAME AND ADDRESS Dept. of Electrical Engineering Rochester Institute of Technology Rochester, New York 14623		10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS 61153N RR021-01 RR021-01-01 NR371-014
11. CONTROLLING OFFICE NAME AND ADDRESS Department of the Navy Office of Naval Research Arlington, Virginia 22217		12. REPORT DATE November 1981
		13. NUMBER OF PAGES 20
14. MONITORING AGENCY NAME & ADDRESS (if different from Controlling Office)		15. SECURITY CLASS. (of this report) UNCLASSIFIED
		15a. DECLASSIFICATION/DOWNGRADING SCHEDULE
16. DISTRIBUTION STATEMENT (of this Report) <div style="border: 1px solid black; padding: 5px; text-align: center;"> DISTRIBUTION STATEMENT A Approved for public release Distribution is unlimited </div>		
17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)		
18. SUPPLEMENTARY NOTES		
19. KEY WORDS (Continue on reverse side if necessary and identify by block number) 1) Approximation Problem 2) Complex Exponentials 3) Pencil of Function 4) Pole-zero Modeling 5) Wiener least squares		
20. ABSTRACT (Continue on reverse side if necessary and identify by block number) The approximation of a function by a sum of complex exponentials, in general, is a nonlinear optimization problem. The optimization problem, however, is linearized through the application of the pencil of function method. This non-iterative method yields the best exponential approximation for a given order of approximation. The method differs radically from the classical Wiener least squares approach in the sense that exponents calculated by the pencil of function method are directly proportional to the integrated squared error in the approximation. As the integrated squared error approaches zero,		

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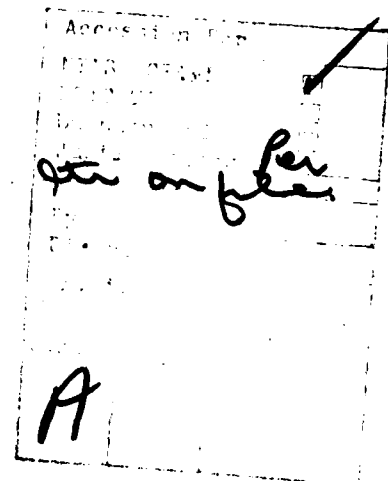
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I. INTRODUCTION

The approximation of an arbitrary real function $x(t)$ which is nonzero for $t \geq 0$, by a sum of complex exponentials often leads to a nonlinear least squares problem. Recently Goulb [1] has presented a class of least squares problems in which the variables separate. One of the objectives of this paper is to show that the pencil of function method also decouples the approximation problem [2-3]. In this presentation we apply the pencil of function method to both the continuous function $x(t)$ and $y(t)$, which is the discretized version of $x(t)$. We show that the resulting expressions are quite different for the two cases. First we present the discrete approximation problem followed by the continuous case. Finally examples are presented to illustrate the stability of this method in the calculation of the exponentials particularly for noisy data.

2. Definition of the Problem

The problem of interest is to approximate an arbitrary square integrable function $x(t)$ by a sum of complex exponentials, i.e.

$$x(t) \approx \sum_{i=1}^{\ell} \sum_{k=1}^{m_i} A_{ik} \cdot \{t\}^{k-1} \cdot \exp\{s_i t\} \quad (1)$$

where

$$\sum_{i=1}^{\ell} m_i = n \quad (2)$$

Here s_i are referred to as the poles of the system and m_i is the order of the pole s_i . n is the total number of poles counting the multiplicities of repeated poles.

In most problems, however, we do not deal with continuous functions but with sampled sequences. So we will assume that the sequence $\{y(p)\}$ has been sampled at a sufficiently fast rate from the continuous function $x(t)$ so that no information has been lost. Thus for a sampled data system we have

$$y(p) = \sum_{i=1}^l \sum_{k=1}^{m_i} A_{ik} \{p\}^{k-1} \{z_i\}^p \quad (3)$$

where

$$z_i = \exp\{s_i T\} \quad (4)$$

where T is the sampling interval. Here z_i are referred to as the z -domain poles.

The problem of approximating A_{ik} and s_i in (1) or A_{ik} and z_i in (3) is a nonlinear approximation problem. However we can linearize this nonlinear problem through the introduction of the T operator.

3. Properties of the T Operator.

The T operator is defined to be an operator similar to a reverse integral operator. It is defined on the space of squared integrable functions in the following way:

$$T[Y] \triangleq T\{y(p)\} \triangleq \left\{ \sum_{k=p}^{\infty} y(k) \right\} \quad (5)$$

so for example

$$T\{0,1,0,0,\dots\} = \{1,1,0,0,0,\dots\} \quad (6)$$

and

$$T\{p(0.5)^p\} = \{2p(0.5)^p + 2(0.5)^p\} \quad (7)$$

From the definition of T in (5) we can prove the following:

(a) T is a linear operator, since

$$T[\alpha Y + \beta Z] = \alpha T[Y] + \beta T[Z] \quad (8)$$

where α and β are scalar quantities. $[Y]$ and $[Z]$ are two sampled data sequences as defined in (5) .

(b) A constant multiplication of a unit impulse remains unchanged by the application of the T operator, i.e.

$$T[a\delta] \triangleq T\{a, 0, 0, \dots\} = aT\{1, 0, 0, \dots\} = aT[\delta] \quad (9)$$

(c) for simple poles (i.e. $m_i=1$ for all i) we have

$$T[Y] = T\{y(p)\} = T\left\{\sum_{i=1}^n A_i (z_i)^p\right\} = \left\{\sum_{i=1}^n A_i \frac{(z_i)^p}{1-z_i}\right\} \quad (10)$$

In particular we have

$$T\{A(z)\} = \left\{\frac{A(z)^p}{1-z}\right\} \quad (11)$$

(d) Successive operations of T are obtained in the following way:

$$T^0 = 1$$

$$T^k[Y] = T^{k-1}[TY] \quad (12)$$

and thus we can generate a polynomial in T. One such polynomial, which will be of fundamental importance, is

$$[1-(1-z)T]^k$$

(e) Also, it can be shown that

$$(1-z)T\{(p)^i(z)^p\} = (p)^i(z)^{p+z} \sum_{k=0}^{i-1} \binom{i}{k} T\{(p)^k(z)^p\} \quad (13)$$

where

$$\binom{i}{k} = \frac{i!}{k!(i-k)!} \quad (14)$$

Properties (c) and (e) demonstrate that the application of the operator T on the sequence $\{y(p)\}$ preserves the poles of the sequence.

4. Properties of the $[1-(1-z)T]^k$ Polynomials:

From property (c) it is clear that if $\{y(p)\}$ be a sequence having a pole z repeated with multiplicity m , then the multiplicity of z in the sequences $[1-(1-z)T]^k\{y(p)\}$ is $m-k$, $k=0, 1, 2, \dots, m-k$ (15)

Furthermore,

$$[1-(1-z)T]^k \{y(p)\} = \{0\} \quad \text{for } k \geq m \quad (16)$$

Geometrically, the above property asserts that the operator $[1-(1-z)T]^k$ projects Y into a lower dimensional hyperplane in z of order $m-k$. Hence, it follows that, if the operator

$$\prod_{i=1}^n [1-(1-z)T]^{m_i} \quad (17)$$

is applied to $[Y]$ which is given by

$$[Y] = \{y(p)\} = \left\{ \sum_{i=1}^n \sum_{k=1}^{m_i} A_{ik}(p)^{k-1} (z_i)^p \right\} \quad (18)$$

then the result would be the null sequence $\{0\}$. We have thus linearized the problem by first attempting to estimate the poles z_i . After the poles z_i have been computed, the residues A_{ij} at the poles can be obtained by a least squares procedure.

5. Determination of the Poles z_i .

In order to determine the poles z_i , we define the following operations. Let

$$[Y_2] \triangleq T[Y_1] \triangleq T[TY_0] = T^2[Y_0] \quad (19)$$

and

$$[Z_1] \triangleq [1-(1-z)T][Y_1] = [1+dT][Y_1] \quad (20)$$

where

$$d \triangleq -(1-z) \quad (21)$$

From (20) it is seen that

$$[Z_1] = [Y_1] + d[T][Y_1] = [T]\{Y_{1-1} + d[T][Y_{1-1}]\} = [T][Z_{1-1}] \quad (22)$$

The sequence $[Z_1]$ is called a pencil of functions parameterized by the scalar parameter d [2-5]. It is a linear combination of functions $[Y_1]$ and $[Y_{1+1}]$ through the parameter d . The pencil of functions contain very important characteristics from a system identification

point of view. When the set $[Z_1]$ of $n+1$ functions becomes linearly dependent then $(d+1)$ becomes a system pole. Now we define the gram matrix $[H]$ whose i th row and j th column are defined as

$$[H] = \left(h_{ij} \triangleq \langle Z_i, Z_j \rangle = \langle [1+dT][Y_i], [1+dT][Y_j] \rangle \right) \\ = [d^* \{ d \langle Y_{i+1}, Y_{j+1} \rangle + \langle Y_i, Y_{j+1} \rangle + d \langle Y_{i+1}, Y_j \rangle + \langle Y_i, Y_j \rangle \}] \quad (23)$$

where

$$\langle Y_i, Y_j \rangle = \sum_{p=0}^{\infty} \{ y_i(p) \} \{ y_j(p) \} \quad (24)$$

and d^* denotes the complex conjugate of d . The underbar distinguishes a matrix from a sequence.

The significance of the gram matrix $[H]$ lies in the fact that the set $[Z_1], [Z_2], [Z_3] \dots [Z_k]$ are linearly independent if and only if $\det[H] > 0$, whereas it is linearly dependent if and only if $\det[H] = 0$. Thus in order that there are n poles in the sequence $[Z_i]$ it is necessary and sufficient that $\det[H]$ be positive or zero according as $k < n$, or $k \geq n$. [This is clear from (16)]

After some very tedious algebraic manipulations it can be shown that

$$\det[H] = \sum_{i=1}^k \sum_{j=1}^k (d^*)^{i-1} (d)^{j-1} Q_{ij} \quad (25)$$

where Q_{ij} are the diagonal cofactors of the matrix $[G]$ whose elements are defined as

$$[G] = [g_{ij} \triangleq \langle Y_i, Y_j \rangle = \sum_{p=0}^{\infty} \{ y_i(p) \} \{ y_j(p) \}] \quad (26)$$

An on line procedure for determining (26) is given by the following equations:

$$x_i(k) = y_i(k) \\ x_i(k) = \sum_{v=k}^{m-1} x_{i-1}(v)$$

$$g_{ij} = (-1)^{i+j} \sum_{k=0}^M [x_i(k)x_j(k) - \sum_{v=1}^i (k-v)^{i-v} x_v(M)x_j(k) - \sum_{\mu=1}^j (k-M)^{j-\mu} x_\mu(M)x_i(k) + \sum_{v=1}^i \sum_{\mu=1}^j x_v(M)x_\mu(M)(k-M)^{i+j-v-\mu}]$$

and

$$x_i(M) = \sum_{v=0}^{M-1} x_{i-1}(v).$$

Here it is assumed that $x_i(p)=0$ for $p>M$. This assumption is reasonable for practical problems.

When the order of the approximation k in (25) becomes equal to the order of the system the sets $[Z_1], [Z_2], \dots [Z_n]$ becomes linearly dependent and hence

$$\det[\underline{H}] = 0 = \sum_{i=1}^k \sum_{j=1}^k (d^*)^{i-1} (d)^{j-1} Q_{ij}. \quad (27)$$

Since

$$Q_{ij}^2 + \det[\underline{G}]_n Q_{ij} Q_{jj} = Q_{ii} Q_{jj} \quad (28)$$

and

$$\det[\underline{G}] = 0 \text{ when } \det[\underline{H}] = 0, \quad (29)$$

(27) can be rewritten as

$$\left\{ \sum_{i=1}^{n+1} \sqrt{Q_{ii}} (d^*)^{n-i+1} \right\} \cdot \left\{ \sum_{j=1}^{n+1} \sqrt{Q_{jj}} (d)^{n-j+1} \right\} = 0 \quad (30)$$

From (30) the poles z_i can be obtained from the solution of the following polynomial equation:

$$\sum_{i=1}^{n+1} \sqrt{Q_{ii}} (-1+z)^{n-j+1} = 0 \quad (31)$$

and the poles s_i are obtained from z_i through the transformation

$$s_i = \frac{1}{T} \ln [z_i]$$

Since the function to be approximated is considered to be real there will also be a complex conjugate set of poles s_i^* as indicated by (30).

Once the poles s_i or z_i are determined the unknown constants A_{ij} can be determined from a least squares procedure.

So far we have done the analysis for discrete systems. However the analysis for the continuous system will be done in section 6.

5. Error Analysis in the Presence of Noise.

When the signal $x(t)$ or the sequence $\{y(p)\}$ does not contain any noise then the poles z_i given by (31) are the exact poles of the system. However if noise is present then they no longer represent the system poles. The objective of this present section is to find the error in the location of the poles due to additive noise.

Let $\{e(p)\}$ be the noisy sequence and let there exist an approximant of $\{e(p)\}$ and call it $\{y(p)\}$, which provides the best exponential approximation of the signal in the given subspace. The assumption made here is that the optimum approximant is unique.

For convenience let us call

$$e(p) = y(p) + \epsilon v(p) \quad (32)$$

We now introduce the error sequence in the normalized form

$$[\{e(p) - y(p)\}] = \epsilon \{r(p)\} \quad (33)$$

where ϵ is a non-negative real number chosen to establish the equality

$$\sum_p \{r(p)\}^2 = \sum_p \{y(p)\}^2 \quad (34)$$

Clearly ϵ represents the fractional error in the representation and is therefore a measure of the degree of approximation of $\{q(p)\}$. If $\{y(p)\}$ is the best approximant of $\{e(p)\}$, then $\{r(p)\}$ must be orthogonal to $\{y(p)\}$. Thus, in particular, $\epsilon=0$ characterizes the perfect representation.

Now consider the following sequences:

$$\begin{aligned}
[E_1] &= \{e_1(p)\} = \{e(p)\} \\
[E_2] &= \{e_2(p)\} = T\{e_1(p)\} \\
&\vdots \\
[E_{k+1}] &= \{e_{k+1}(p)\} = T\{e_k(p)\} = T\{y_k(p) + \epsilon r_k(p)\}
\end{aligned} \tag{35}$$

In (35) $\{y(p)\}$ and $\{r(p)\}$ are not known explicitly, nor is ϵ for that matter. We begin with the computation of the $(n+1) \times (n+1)$ gram matrix $[B]$ for the sequence $[E_1], [E_2], \dots, [E_n]$, i.e.

$$\begin{aligned}
[B] &= [b_{ij}] \triangleq \langle \{e_i(p)\}, \{e_j(p)\} \rangle \\
&= \begin{Bmatrix} y_1(p) + \epsilon r_1(p) \\ \vdots \\ y_{n+1}(p) + \epsilon r_{n+1}(p) \end{Bmatrix} \cdot \begin{Bmatrix} y_1(p) + \epsilon r_1(p) \dots y_{n+1}(p) + \epsilon r_{n+1}(p) \end{Bmatrix}
\end{aligned} \tag{36}$$

Since $\{y(p)\}$ is a sequence of order n , the corresponding $[Y_1], \dots, [Y_{n+1}]$ sequences are linearly dependent. Thus there exists then a nonsingular matrix $[P]$ such that

$$[Y] \triangleq \begin{Bmatrix} \{y_1(p)\} \\ \vdots \\ \{y_{n+1}(p)\} \end{Bmatrix} = [P] \begin{Bmatrix} \{\bar{y}_1\} \\ \vdots \\ \{\bar{y}_n\} \\ 0 \end{Bmatrix} = [P][\bar{Y}] \tag{37}$$

From the theory of orthogonal transformation the sequences $\{\bar{y}_1(p)\}, \{\bar{y}_2(p)\}, \dots, \{\bar{y}_n(p)\}$ must form an orthonormal set.

Similarly we obtain

$$[R] = \begin{Bmatrix} \{r_1(p)\} \\ \vdots \\ \{r_{n+1}(p)\} \end{Bmatrix} = [P] \begin{Bmatrix} \{\bar{r}_1(p)\} \\ \vdots \\ \{\bar{r}_{n+1}(p)\} \end{Bmatrix} = [P][\bar{R}] \tag{38}$$

Observe in (38) the sequences $\{\bar{r}_1(p)\}, \dots, \{\bar{r}_{n+1}(p)\}$ are not orthonormal to each other. Substitution of (37) and (38) into (36) yields

$$\begin{aligned}
[\underline{B}] &= [\underline{P}] [\underline{Y} + \epsilon \underline{R}] [\underline{Y} + \epsilon \underline{R}]^T [\underline{P}]^T \\
&= [\underline{P}] \{ [\underline{Y}] [\underline{Y}]^T + \epsilon [\underline{R}] [\underline{Y}]^T + \epsilon [\underline{Y}] [\underline{R}]^T + \epsilon^2 [\underline{R}] [\underline{R}]^T \} [\underline{P}]^T
\end{aligned} \quad (39)$$

Observe that

$$[\underline{Y}] [\underline{Y}]^T = \begin{bmatrix} I_{n \times n} & \begin{matrix} 0 \\ \vdots \\ 0 \end{matrix} \\ \begin{matrix} 0 \\ \vdots \\ 0 \end{matrix} & \begin{matrix} 0 \\ \vdots \\ 0 \end{matrix} \end{bmatrix}_{(n+1) \times (n+1)} \quad (40)$$

where I is the $n \times n$ identity matrix and

$$[\underline{R}] [\underline{Y}]^T = \begin{Bmatrix} \{\bar{r}_1(p)\} \cdot \{\bar{y}_1(p)\} \dots \{\bar{r}_1(p)\} \cdot \{\bar{y}_n(p)\} \\ \{\bar{r}_{n+1}(p)\} \cdot \{\bar{y}_1(p)\} \dots \{\bar{r}_{n+1}(p)\} \cdot \{\bar{y}_n(p)\} \end{Bmatrix} \quad (41)$$

By utilizing the theorem on determinant expansion as presented in Appendix A, we get

$$\begin{aligned}
\det[\underline{B}] &= (\det[\underline{P}])^2 \cdot \det \left[\begin{bmatrix} I_{n \times n} & \begin{matrix} \epsilon \{\bar{r}_{n+1}(p)\} \cdot \{\bar{y}_1(p)\} \\ \vdots \\ \epsilon \{\bar{r}_{n+1}(p)\} \cdot \{\bar{y}_n(p)\} \end{matrix} \\ \begin{matrix} 0 \\ \vdots \\ 0 \end{matrix} & \begin{matrix} \epsilon^2 \{\bar{r}_{n+1}(p)\} \cdot \{\bar{y}_{n+1}(p)\} \\ \vdots \\ \epsilon^2 \{\bar{r}_{n+1}(p)\} \cdot \{\bar{y}_{n+1}(p)\} \end{matrix} \end{bmatrix} \right] \\
&\quad + \text{terms of the order of } \epsilon^3 \text{ and higher} \\
&= \epsilon^2 \cdot \|\{\bar{r}_{n+1}(p)\}\|^2 \cdot (\det[\underline{P}])^2 + O(\epsilon^3) + \dots
\end{aligned} \quad (42)$$

Thus it is clear that $\sqrt{\det[\underline{B}]}$ is of the same order of magnitude as the error measure ϵ , i.e.

$$\det[\underline{B}] = O(\epsilon) \quad (43)$$

In particular, a necessary and sufficient condition for perfect approximation of a signal of order n is that $\det[\underline{B}]$ must vanish.

In order to find the error in the location of the poles first the error in the diagonal cofactors of the gram matrix $[\underline{B}]$ has to be estimated. The diagonal cofactors are given by:

$$D_{ii} = \det \left\{ \begin{matrix} \{y_1(p) + \epsilon r_1(p)\} \\ \vdots \\ \{y_{i-1}(p) + \epsilon r_{i-1}(p)\} \\ \{y_{i+1}(p) + \epsilon r_{i+1}(p)\} \\ \vdots \\ \{y_{n+1}(p) + \epsilon r_{n+1}(p)\} \end{matrix} \right\} \cdot \left\{ \begin{matrix} \{y_1(p) + \epsilon r_1(p)\} \dots \{y_{i-1}(p) + \epsilon r_{i-1}(p)\} \\ \vdots \\ \{y_{i+1}(p) + \epsilon r_{i+1}(p)\} \dots \{y_{n+1}(p) + \epsilon r_{n+1}(p)\} \end{matrix} \right\} \quad (44)$$

Let $[\underline{S}_i]$ be the matrix which orthonormalizes the vectors, from $\{y_1(p)\} \dots \{y_{i-1}(p)\} \{y_i(p)\} \dots \{y_{n+1}(p)\}$ to $\{\bar{y}_1(p)\} \dots \{\bar{y}_{i-1}(p)\} \{\bar{y}_{i+1}(p)\} \dots \{\bar{y}_{n+1}(p)\}$. So we have

$$\left\{ \begin{matrix} \{y_1(p)\} \\ \vdots \\ \{y_{i-1}(p)\} \\ \{y_{i+1}(p)\} \\ \vdots \\ \{y_{n+1}(p)\} \end{matrix} \right\} = [\underline{S}_i] \cdot \left\{ \begin{matrix} \{\bar{y}_1(p)\} \\ \vdots \\ \{\bar{y}_{i-1}(p)\} \\ \{\bar{y}_{i+1}(p)\} \\ \vdots \\ \{\bar{y}_{n+1}(p)\} \end{matrix} \right\} \triangleq [\underline{S}_i][\bar{\underline{Y}}_i] \quad (45)$$

$$\left\{ \begin{matrix} \{r_1(p)\} \\ \vdots \\ \{r_{i-1}(p)\} \\ \{r_{i+1}(p)\} \\ \vdots \\ \{r_{n+1}(p)\} \end{matrix} \right\} = [\underline{S}_i] \cdot \left\{ \begin{matrix} \{\bar{r}_1(p)\} \\ \vdots \\ \{\bar{r}_{i-1}(p)\} \\ \{\bar{r}_{i+1}(p)\} \\ \vdots \\ \{\bar{r}_{n+1}(p)\} \end{matrix} \right\} \triangleq [\underline{S}_i][\bar{\underline{R}}_i] \quad (46)$$

Observe that the vectors $\{\bar{r}_1(p)\} \dots \{\bar{r}_{n+1}(p)\}$ are not a set of orthonormal vectors. Substitution of (45) and (46) into (44) yields

$$\begin{aligned} D_{ii} &= \det \left[[\underline{S}_i][\bar{\underline{Y}}_i + \epsilon \bar{\underline{R}}_i][\bar{\underline{Y}}_i + \epsilon \bar{\underline{R}}_i]^T [\underline{S}_i]^T \right] \\ &= (\det[\underline{S}_i])^2 \cdot \det \left[[\bar{\underline{Y}}_i][\bar{\underline{Y}}_i]^T + \epsilon [\bar{\underline{R}}_i][\bar{\underline{Y}}_i]^T + \epsilon [\bar{\underline{Y}}_i][\bar{\underline{R}}_i]^T + \epsilon^2 [\bar{\underline{R}}_i][\bar{\underline{R}}_i]^T \right] \quad (47) \end{aligned}$$

By utilizing

$$[\bar{Y}_1][\bar{Y}_1]^T = [I]_{n \times n}$$

and applying the theorem on determinant expansion [as presented in Appendix A] we obtain

$$D_{ii} = Q_{ii} + 2\epsilon(\det[S_i])^2 \cdot \left\{ \sum_{j=1}^{n+1} \sum_{p=0}^{\infty} \{\bar{y}_j(p)\} \cdot \{r_j(p)\} \right\} \quad (48)$$

where Q_{ii} are the diagonal cofactors of the gram matrix formed by the sets $[Y_1], [Y_2] \dots [Y_{n+1}]$ and is given by (26). The underlying assumption in (48) is of course we have the right order of the system n , i.e.

$$\det[G] = 0. \quad (49)$$

Observe that (48) implies continuous differentiability of the diagonal cofactors D_{ii} with respect to the error ϵ , so that

$$\lim_{\epsilon \rightarrow 0} D_{ii} = Q_{ii} \text{ for all } i. \quad (50)$$

Let z'_i be the approximate poles that are obtained by utilizing the diagonal cofactors D_{ii} instead of Q_{ii} in (31). So the poles z'_i are the roots of the following equation

$$\sum_{i=1}^{n+1} \sqrt{D_{ii}} (-1+z'_i)^{n+1-i} = 0 \quad (51)$$

Substitution of (48) into (51) illustrates that the approximate poles z'_i are related to the exact poles z_i by the relationship

$$z'_i = z_i + \epsilon F_i + \Theta(\epsilon^2) \quad (52)$$

where F_i are certain constants. The approximate poles z'_i are again a continuous function of ϵ . So the error in the approximation $(z'_i - z_i)$ is a continuous function of ϵ and approaches zero uniformly as $\epsilon \rightarrow 0$.

This is the unique feature of this technique. The continuous dependence of the error in the pole locations on ϵ is not guaranteed by other optimization techniques and the classical nonlinear methods based on Wiener filter theory [4-5].

6. Continuous Version of the Pencil of Function Method.

In the previous sections the approximation problem involved discretizing a continuous function which was then processed digitally. A strictly continuous version is also available. The problem is to approximate the continuous function $x(t)$ in the form given by (1), i.e.

$$x(t) = \sum_{i=1}^{\ell} \sum_{k=1}^{m_i} A_{ik} \{t\}^{k-1} \exp\{s_i t\}$$

In this case we want to estimate s_i instead of $\exp\{s_i T\}$. For the continuous case the T operator of the previous sections is replaced by the reverse time integral operator so that

$$x_{i+1}(t) = \int_t^{\infty} x_i(f) df \quad (53)$$

and $x_1(t) = x(t)$

The two sided Laplace Transform of (53) is given by

$$X_{i+1}(s) = \int_{-\infty}^{\infty} X_{i+1}(t) e^{-st} dt = \frac{X_i(s)}{s} \quad (54)$$

and

$$X_i(s) = \sum_{j=1}^{\ell} \sum_{k=1}^{m_j} \frac{A_{jk}}{s^k} \frac{(j-k)!}{(s-s_i)^j} = \frac{N(s)}{\prod_{i=1}^{\ell} (s-s_i)^{m_i}} \quad (55)$$

Next we consider the pencil of functions defined by

$$\{x_1(t) - \lambda x_2(t)\}, \{x_2(t) - \lambda x_3(t)\}, \dots, \{x_n(t) - \lambda x_{n+1}(t)\} \quad (56)$$

Since for this problem it is simpler to deal with the Laplace transform, attention is focused on the set of pencil of functions defined by

$$\{X_1(s) - \lambda X_2(s)\}, \{X_2(s) - \lambda X_3(s)\}, \dots, \{X_n(s) - \lambda X_{n+1}(s)\} \quad (57)$$

The above set of pencil of functions displays some interesting properties if checked for linear independence. Checking for linear independence, we write

$$\sum_{k=1}^n [a_k \{X_k(s) - \lambda X_{k+1}(s)\}] = 0 \quad (58)$$

Substitution of (54) into (58) yields

$$\sum_{k=1}^n s^{n-k-1} [a_k (s-\lambda) X(s)] = 0$$

or

$$\sum_{k=1}^n s^{n-k} [a_k (s-\lambda) \frac{N(s)}{\prod_{i=1}^m (s-s_i)^{m_i}}] = 0 \quad (59)$$

Equation (59) represents a polynomial in s of the order n . Since $1, s, s^2, \dots, s^{n-1}$ span a n dimensional subspace, this set is linearly independent and hence the set of n coefficients a_k must be identically zero if (59) is to be satisfied for all values of s . The above holds as long as $\lambda \neq \lambda_i$. For $\lambda = \lambda_i$, it turns out that (59) has a root $s = \lambda_i$ which can be factored out. The set $1, s, s^2, \dots, s^{n-2}$ spans a $(n-1)$ dimensional subspace. Nevertheless there remains a set of coefficients a_k . It is now possible for (59) to be satisfied with at least one non-zero coefficient. This implies linear dependence of the set (57) when $\lambda = \lambda_i$. So if λ is a system pole then the gram determinant of the set of functions in (57) will be zero. The gram determinant of the given set in (56) is defined by

$$[F] = \begin{bmatrix} \langle x_1 - \lambda x_2, x_1 - \lambda x_2 \rangle & \dots & \langle x_1 - \lambda x_2, x_n - \lambda x_{n+1} \rangle \\ \vdots & \dots & \vdots \\ \langle x_n - \lambda x_{n+1}, x_1 - \lambda x_2 \rangle & \dots & \langle x_n - \lambda x_{n+1}, x_n - \lambda x_{n+1} \rangle \end{bmatrix} \quad (60)$$

where

$$\langle x_i, x_j \rangle = \int_0^{\infty} x_i(t) x_j(t) dt \quad (61)$$

Thus $\det[F] = 0$ if λ is a system pole. After some tedious algebraic manipulations it can be shown that

$$\det[F] = \left\{ \sum_{i=1}^{n+1} \sqrt{\Delta_{ii}} \lambda^{n-i+1} \right\} \left\{ \sum_{j=1}^{n+1} \sqrt{\Delta_{jj}} \lambda^{n-j+1} \right\} = 0 \quad (62)$$

where Δ_{kk} is the k th diagonal cofactor in the grammian $[F^1]$ which is defined by

$$[F^1] = \begin{bmatrix} \langle x_1, x_1 \rangle & \dots & \langle x_1, x_{n+1} \rangle \\ \vdots & \dots & \vdots \\ \langle x_{n+1}, x_1 \rangle & \dots & \langle x_{n+1}, x_{n+1} \rangle \end{bmatrix} \quad (63)$$

The polynomial equation whose roots are the pole locations, is then given by

$$\sum_{i=1}^{n+1} \sqrt{\Delta_{ii}} \lambda^{n-i+1} = 0 \quad (64)$$

After the poles s_i have been obtained the residues are obtained by a least squares fashion.

7. Examples

As an example consider the transient response of a conducting pipe tested at the ATHAMAS-I Electromagnetic Pulse (EMP) simulator. The conducting pipe is 10m long and 1m in diameter. Hence the true resonance of the pipe is expected to be in the neighborhood of 14MHz. Also the pipe has been excited in such a way that it is reasonable to expect only odd harmonics at the scattered fields. The data which have been measured are the integral of the electric field and hence is available in terms of a voltage. Thus, in addition to the frequencies of the conducting pipe one should also observe a very dominant low frequency pole. The same transient data as depicted in Figure 1 is used for

analysis. The results for a fifth and a seventh order system are as follows:

For $n = 5$, the poles in radians/sec are

$$\begin{aligned} (-0.0029 \pm j0.083) \times 10^9 & \quad (=13.33\text{MHz}) \\ (-0.0428 \pm j0.217) \times 10^9 & \quad (=35.20 \text{ MHz}) \\ (-0.0098 \quad) \times 10^9 & \quad (= 1.56 \text{ MHz}) \end{aligned}$$

For $n = 7$, the poles in radians/sec are

$$\begin{aligned} (-0.0058 \pm j0.084) \times 10^9 & \quad (=13.40 \text{ MHz}) \\ (-0.0270 \pm j0.219) \times 10^9 & \quad (=35.10 \text{ MHz}) \\ (-0.0270 \pm j0.550) \times 10^9 & \quad (=87.60 \text{ MHz}) \\ (-0.0012 \quad) \times 10^9 & \quad (= 0.19 \text{ MHz}) \end{aligned}$$

It is interesting to observe that the real pole due to the integrator has been obtained. This pole is a very dominant pole as the data were recorded after having passed through an integrator. The above results display a dynamic range of approximately 1000:1 for the values of poles of the conducting pipe.

Next the data were differentiated to get rid of the undesirable dominang pole of the integrator. The differentiation was done numerically. For a fourth and sixth order system the above results have been recalculated as follows:

For $n = 4$, the poles in radians/sec are

$$\begin{aligned} (-0.0026 \pm j0.086) \times 10^9 & \quad (=13.70 \text{ MHz}) \\ (-0.0480 \pm j0.235) \times 10^9 & \quad (=37.47 \text{ MHz}) \end{aligned}$$

For $n = 6$, the poles in radians/sec are

$$\begin{aligned} (-0.005 \pm j0.083) \times 10^9 & \quad (=13.23 \text{ MHz}) \\ (-0.034 \pm j0.221) \times 10^9 & \quad (=35.59 \text{ MHz}) \\ (-0.071 \pm j0.406) \times 10^9 & \quad (= 65.9 \text{ MHz}) \end{aligned}$$

Here a good approximation to the poles has been obtained with only four poles. Also, there seems to be a good agreement in the pole locations obtained from the original integrated data and the numerically differentiated data. It is also interesting to observe that indeed the poles are occurring approximately at odd harmonics of the fundamental.

In figure 2, the true numerically differentiated data are plotted against the reconstructed response of a sixth order system. The plot has been normalized to unity amplitude. It is interesting to note that there is a close agreement even in the very early times of the two waveforms.

8. Conclusion

The pencil of function method has been applied to the approximation of an arbitrary function by a sum of complex exponentials. The important features of this technique are the natural insensitivity to noise and the continuous dependence of the pole locations to the integrated square error. Finally examples have been presented to illustrate the stability of the pole locations yielded by the pencil of function method.

9. References

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10. Appendix

Let A and B be two square NXN matrices. Let the columns of A and B be represented by A_i and B_i for $i=1,2,\dots,N$. Denote

$$\begin{aligned} C_{v_i}^1 &= A_i \quad \text{if } v_i=1 \\ &= B_i \quad \text{if } v_i=0 \end{aligned}$$

and the matrix constituted by the columns of

$$C_{v_1}^1, C_{v_2}^2, C_{v_3}^3, \dots, C_{v_N}^N \quad \text{as} \quad \underline{C}_{v_1, v_2, \dots, v_N}.$$

Then

$$\det |[\underline{A}+\underline{B}]| = \sum_{m=0}^N \det |\underline{C}_{v_1, v_2, \dots, v_N}|$$

where in the second summation exactly m of the v_i 's equal 1 and the rest equal 0.

This is a standard result of determinant expansion.

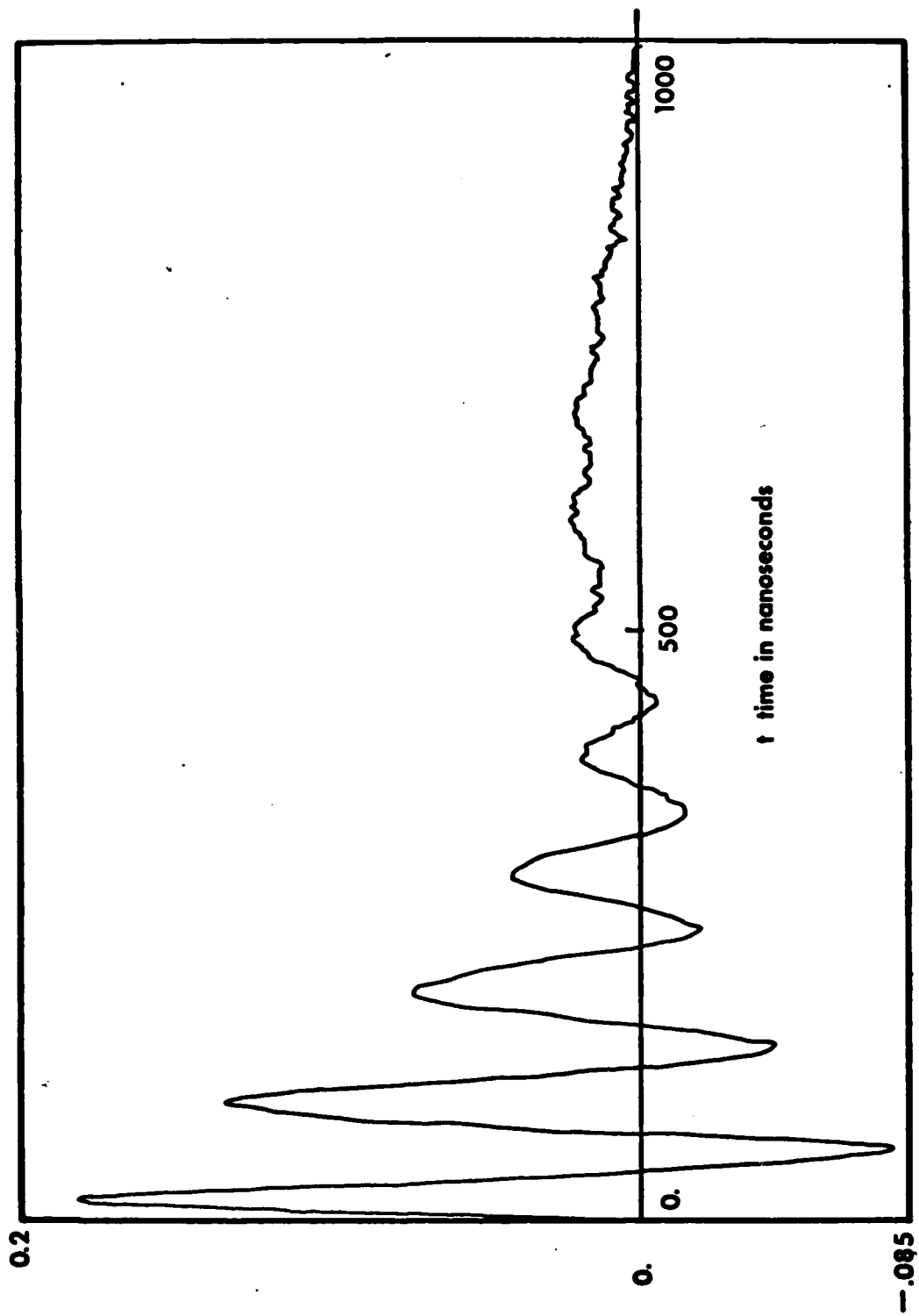


Figure 1. Transient Response of a Conducting Pipe, Measured at the ATHAMAS-1 EMP Simulator.

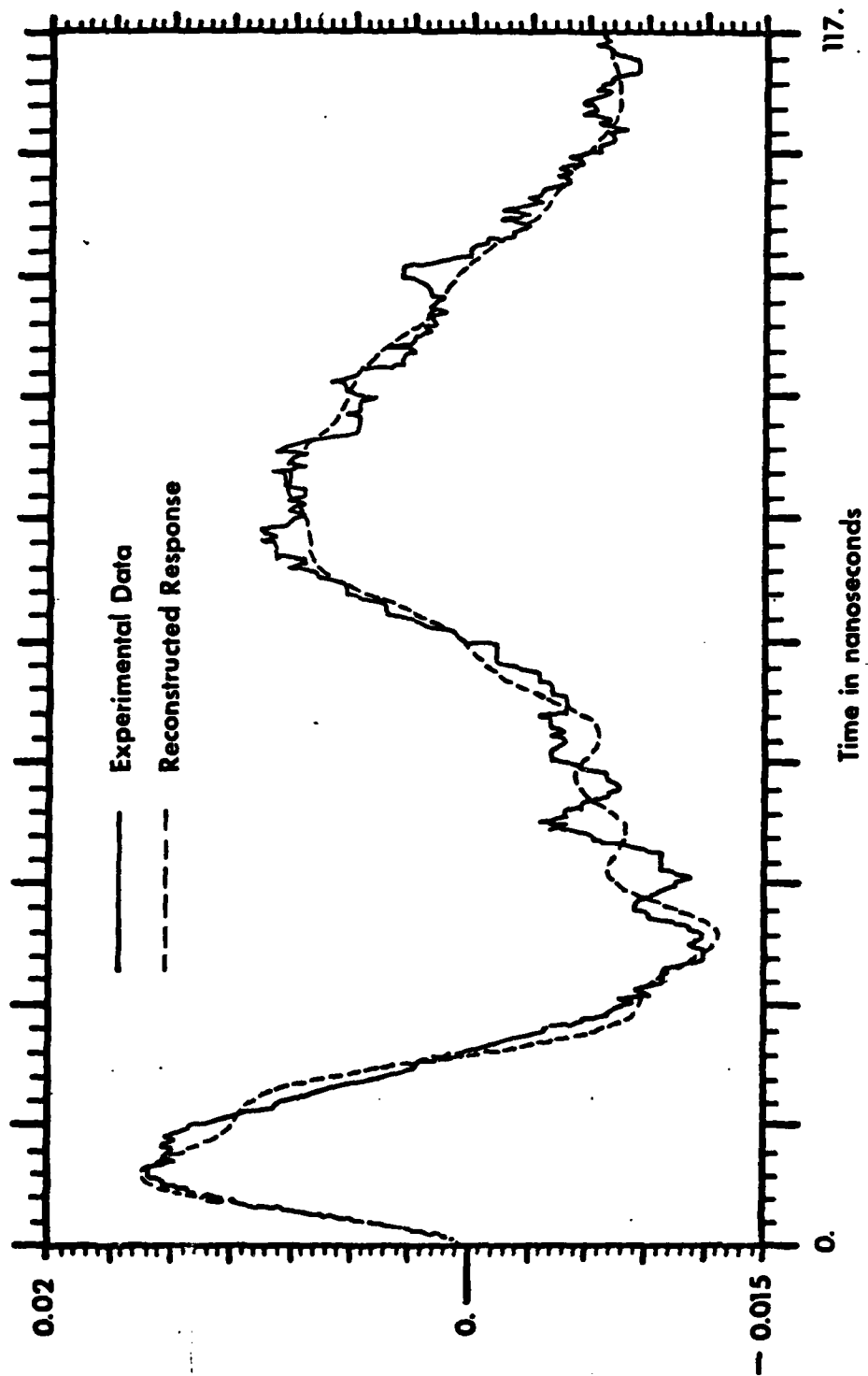


Figure 2. True Response Vs. Reconstructed Response of a Sixth Order System for a 10m long 1m Diameter Conducting Pipe.

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